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RECEPTION OF MULTICARRIER SPREAD-SPECTRUM SIGNALS

Field of the Invention

This invention relates to multicarrier wireless communication systems, and more specifically Orthogonal Frequency Division Multiplex (OFDM) modulation schemes.

Background of the Invention

Such modulation schemes are now widely used in standards as a means to provide high data rates for communication systems including wireless local area networks (WLANs): 'IEEE 802.11a' in USA and 'HIPERLAN/2' in Europe, ADSL (Asynchromous Digital Subscriber Line) over twisted pairs and 'HomePLUG' on powerlines.

For the next decade, the challenge is to deliver an increased data rate coping with the requirements of multimedia broadband transmissions. None of the existing standards will be able to meet these requirements on a larger scale (involving many users) which motivates the search for more robust yet simple modulation schemes that, combined with an appropriate decoding algorithm, show better performance in terms of Packet Error Rate (PER) than classical OFDM systems. This technical criterion translates directly into increased system throughput. Clearly, an attractive property for such a

new modulation scheme would be for it to be viewed as a simple extension of OFDM so that it could be implemented in existing standards as a proprietary transmission mode. In this way it could also provide a means for smooth transition to new standards.

In the field of this invention, enhancements have been proposed as a workaround for alleviating an inherent OFDM weakness: when a carrier is subject to a strong channel attenuation, even in absence of noise, the data conveyed is irremediably lost. The classical alternative is to use forward error correction (FEC) coding to spread the information along the carriers, but another strategy has been proposed: to combine the strength of OFDM and CDMA by pre-processing the block of symbols to be transmitted by a unitary spreading matrix $\mathbf{W}$ (often chosen to be a Walsh Hadamard transform for its attractive implementation properties) prior to the FFT/IFFT (Fast Fourier Transform/Inverse FFT) modulation.

This redundantless precoder $\mathbf{W}$ has the role of uniformly spreading the information to be transmitted on all the carriers so that even if one carrier is unrecoverable, the information transmitted can still be retrieved by decoding of other subbands.

Implementations of such spread OFDM (SOFDM also known as single user OFDM-CDMA with cyclic prefix) modulation techniques require successive interference cancellation (SIC), and many SIC algorithms have been proposed. One of the most well known is 'V-BLAST' proposed by Bell Labs

for multiple antennas systems in the publication by G.J. Foschini and M.J. Gans, "On Limits of Wireless Communications in a fading Environment when Using Multiple Antennas", *Wireless Personnal Communications* 6:311-335, 1998. However, it has been demonstrated (in the publication by P. Loubaton, M. Debbah and M. de Courville, "Spread OFDM Performance with MMSE Equalization", in *International Conference on Acoustics, Speech, and Signal Processing*, Salt Lake City, USA, May 2001) that V-BLAST algorithms are not suited for conventional SOFDM systems due to the averaging of the SNRs (signal/noise ratios) at the receiver across the carriers during the despreading step. Moreover, such approaches lead to a tremendous decoding complexity due to the computation of several pseudo inverse matrices.

A need therefore exists for an OFDM communication system and decoding algorithm for use therein wherein the abovementioned disadvantage(s) may be alleviated.

Statement of Invention

The present invention provides a method of decoding a received spread OFDM wireless communication signal, and a receiver comprising decoding means for decoding a received signal by such a method, in a spread OFDM wireless communication receiver, as described in the accompanying claims.

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In one embodiment of the present invention, the decoding algorithm comprises splitting a received block into two equal parts, one of the parts being decoded first and then subtracted from the received vector to suppress part of the interference and the other of the parts being decoded next. This iterative procedure can be further extended by successive block splitting and results in a multi-resolution decoding algorithm. An attractive property of this algorithm is that although it still relies on the computation of pseudo-inverses, the expressions of these pseudo-inverses are easy to derive and may consist simply in the product of a diagonal matrix by a Walsh Hadamard transform. Thus, using Walsh Hadamard spreading sequences, the inherent complexity penalty of a V-BLAST decoding schemes is simply removed. This allows a significant gain in performance (e.g., around 3-4dB compared to MMSE SOFDM) with only a modest increase in complexity, which motivates:

i) the use of such new modulation schemes in practice and

ii) their proposal as a solution for future wireless LAN standards.

The following technical merits of the multi-resolution decoding algorithm of this embodiment of the present invention can be highlighted:

- Low arithmetical complexity compared to existing SIC BLAST techniques with same or better performance.
- Flexibility and scalability of the method (it is possible to adjust the number of iterations to be

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performed based on a performance/complexity tradeoff).

- Can be combined into all OFDM standards as a proprietary transmission mode (since it can be

viewed as a simple extension of current OFDM systems)

- Yields a significant PER performance enhancement compared to classical OFDM and minimum mean square error (MMSE) SOFDM receivers (e.g., 3dB).

Brief Description of the Drawings

One OFDM single user communication system and decoding algorithm for use therein incorporating the present invention will now be described, by way of example only, with reference to the accompanying drawing(s), in which:

FIG. 1 shows a block schematic diagram of a OFDM-CDMA (spread OFDM) single user communication system;

FIG. 2 shows a blcok schematic representation of the system of FIG. 1 modeled in the frequency domain;

FIG. 3 shows a diagrammatic binary tree representation of the two-stage multi-resolution decoding algorithm used in the system of FIG. 1;

FIG. 4 and FIG. 5 show graphical representations of simulation performance of the multi-resolution decoding algorithm compared with other decoding scenarios under different respective channel profiles in terms of BER (bit error rate) as a function of $\frac{E_b}{N_0}$ (energy per bit / noise energy).

Description of Preferred Embodiment(s)

As will be explained below, the decoding algorithm to be described significantly enhances performance compared to MMSE equalized SOFDM scheme, with a complexity excess that is marginal compared to V-BLAST decoding strategies.

Consider the dimension $N\times 1$ vector s representing the block of complex valued symbols to be transmitted (each one belonging to a finite alphabet called constellation, e.g., QPSK, QAM, etc.). The overall Spread-OFDM transmission system of interest 100 depicted in FIG. 1 includes, in a transmitter, a spreading matrix module 110, a module 120 providing modulation, a module 130 providing guard interval insertion and parallel-to-serial conversion, and a digital-to-analog converter 140. The transmitter is coupled via a wireless communication channel 150 to a receiver including a mixer and analog-to-digital converter 160, a module 170 providing guard interval suppression and serial-to-parallel conversion, a module 180 providing demodulation, and a module 190 providing demodulation.

The system of FIG. 1 can be modelled directly in the frequency domain as illustrated in FIG. 2 so that the received vector y expresses as:

$$y = HWs + b = Ms + b$$

where:

H is a $N \times N$ diagonal matrix, bearing the complex frequency channel attenuations,

W is a $N \times N$ unitary Walsh Hadamard spreading matrix, whose particular recursive structure is exploited in the decoding algorithm to reduce complexity,

b is a $N \times 1$ complex white IID (independent and identically distributed) Gaussian noise vector whose component variance is $E\left[|b_k|^2\right] = \sigma^2$ (Ere presenting the mathematical expectation operator).

In the following anaysis, H, W and σ^2 are assumed to be known at the receiver by any given classical estimation technique.

The procedure described below deals with the retrieval of the information vector s based on the received vector y which is referred as the equalization step. Instead of using a traditional MMSE equalizer, a specific successive interference cancellation algorithm (termed a 'multi-resolution decoding algorithm') will be described. In the following analysis, $()^h$ is defined as the Hermitian transpose operator and I_N is defined as the NxN identity matrix.

The multi-resolution decoding algorithm is based on the following steps:

(i) Decode the received y vector by an MMSE equalizer followed by a non-linear decision

function denoted by *dec()* (e.g., hard decision demapper, soft decision, etc.) $\hat{s}=dec(G_{MMSE}\,y)$ where $G_{MMSE}=M^h(M\,M^h+\sigma^2 I_N)^{-1}$ (convenient implementations of the product $G_{MMSE}\,y$ are detailed below).

(ii) Split the vector $\hat{s}$ in two equal size $N/2$ parts $\hat{s}=\begin{bmatrix}\hat{s}_1\\ \hat{s}_2\end{bmatrix}$.

(iii) Subtract the second half $\hat{s}_2$ of the vector $\hat{s}$ from the received vector y to remove the interference generated by the first half of s (treating s_2 as if $\hat{s}_2=s_2$.

(iv) Perform an MMSE equalization of the resulting y_1 half-sized vector by matrix G_1, followed by the decision function *dec()* for obtaining a more reliable estimate $\hat{\hat{s}}_1$ of s_1 than $\hat{s}_1$.

(v) Possibly reiterate the procedure, this time on the first half of $\hat{s}$ for retrieving a better estimate $\hat{\hat{s}}_2$ of s_2 than $\hat{s}_2$.

(vi) These operations can be repeated substituting $\hat{s}_1$ and $\hat{s}_2$ by $\hat{\hat{s}}_1$ and $\hat{\hat{s}}_2$ respectively

Translated into equations, this amounts to the following steps:

First stage (310) of the multi-resolution decoding algorithm

Step 0, (300) : MMSE equalization of $y : \hat{s} = \begin{bmatrix} \hat{s}_1 \\ \hat{s}_2 \end{bmatrix} = dec(G_{MMSE}$

$y)$

Step 1 : $y_1 := y - M\begin{bmatrix} 0 \\ \hat{s}_2 \end{bmatrix} = M\begin{bmatrix} s_1 \\ s_2 - \hat{s}_2 \end{bmatrix} + b$

Step 2 : MMSE equalization of $\hat{\hat{s}}_1$: $\hat{\hat{s}}_1 = dec(G_1^{(1)} y_1)$

Step 3 : $y_2 := y - M\begin{bmatrix} \hat{\hat{s}}_1 \\ 0 \end{bmatrix} = M\begin{bmatrix} s_1 - \hat{\hat{s}}_1 \\ s_2 \end{bmatrix} + b$

Step 4 : MMSE equalization of $\hat{\hat{s}}_2$: $\hat{\hat{s}}_2 = dec(G_2^{(1)} y_2)$

Step 5 : $\hat{s} = \begin{bmatrix} \hat{s}_1 \\ \hat{s}_2 \end{bmatrix} := \begin{bmatrix} \hat{\hat{s}}_1 \\ \hat{\hat{s}}_2 \end{bmatrix}$

Step 6 : go to Step 1

It should be noted that although as stated above only a subdivision by two of the received vector y is performed, in its more generalized form the procedure can apply to smaller subdivisions of y of length N divided by a power of 2: $N/2^k$ for any integer k such that the result remains an integer. The generalized algorithm consists in reiterating the procedure already explained to each resulting sub-block of y. Let stage i of the algorithm define the operations performed for a level of subdivisions of y in blocks of size $N/2^i$.

As an illustration, the second stage of the proposed multi-resolution algorithm results in the following operations:

Second stage (320) of the multi-resolution decoding algorithm

Step 0: form $\hat{s} = \begin{bmatrix} \hat{s}_1 \\ \hat{s}_2 \\ \hat{s}_3 \\ \hat{s}_4 \end{bmatrix}$

Step 1 : $y_1 := y - M\begin{bmatrix} 0 \\ \hat{s}_2 \\ \hat{s}_3 \\ \hat{s}_4 \end{bmatrix} = M\begin{bmatrix} s_1 \\ s_2 - \hat{s}_2 \\ s_3 - \hat{s}_3 \\ s_4 - \hat{s}_4 \end{bmatrix} + b$

Step 2 : MMSE equalization of $\hat{\hat{s}}_1$: $\hat{\hat{s}}_1 = dec(G_1^{(2)} y_1)$

Step 3 : $y_2 := y - M\begin{bmatrix} \hat{\hat{s}}_1 \\ 0 \\ \hat{s}_3 \\ \hat{s}_4 \end{bmatrix} = M\begin{bmatrix} s_1 - \hat{\hat{s}}_1 \\ s_2 \\ s_3 - \hat{s}_3 \\ s_4 - \hat{s}_4 \end{bmatrix} + b$

Step 4 : MMSE equalization of $\hat{\hat{s}}_2$: $\hat{\hat{s}}_2 = dec(G_2^{(2)} y_2)$

Step 5 : $y_3 := y - M\begin{bmatrix} \hat{\hat{s}}_1 \\ \hat{\hat{s}}_2 \\ 0 \\ \hat{s}_4 \end{bmatrix} = M\begin{bmatrix} s_1 - \hat{\hat{s}}_1 \\ s_2 - \hat{\hat{s}}_2 \\ s_3 \\ s_4 - \hat{s}_4 \end{bmatrix} + b$

Step 6 : MMSE equalization of $\hat{\hat{s}}_3$: $\hat{\hat{s}}_3 = dec(G_3^{(2)} y_3)$

Step 7 : $y_4 := y - M\begin{bmatrix} \hat{\hat{s}}_1 \\ \hat{\hat{s}}_2 \\ \hat{\hat{s}}_3 \\ 0 \end{bmatrix} = M\begin{bmatrix} s_1 - \hat{\hat{s}}_1 \\ s_2 - \hat{\hat{s}}_2 \\ s_3 - \hat{\hat{s}}_3 \\ s_4 \end{bmatrix} + b$

Step 8 : MMSE equalization of $\hat{\hat{s}}_4$: $\hat{\hat{s}}_4 = dec(G_4^{(2)} y_4)$

Step 9 : $\hat{s} := \hat{\hat{s}}$

Step 10 : go to Step 1

Note that $G_l^{(\gamma)}$ denotes the MMSE equalizer matrix at stage γ for the sub-block r of vector y of size $N/2^{\gamma}$.

It is important to note that each stage can be sequenced in many ways following the graphic illustration of FIG. 3 using a binary tree. Each path in the binary tree results into another instantiation of the proposed algorithm. The depth in terms of number of stages and the number of times each of the stages has to be iterated can be determined by a complexity/performance trade-off criterion.

Thus in order to refine the decoding, the same mechanisms can be applied to blocks of size $N/4$, and then $N/8$, etc. leading to a higher resolution of the decoding.

Clearly, increasing the number of stages and iterations yields a more robust estimation procedure. However, simulations show that the bit error rate converges after a few iterations, so to improve again the decoded vector, fortunately in practice only the second stage of the algorithm needs to be considered.

A fast algorithm for computing the product of vector y_l by matrix $G_l^{(\gamma)}$ be implemented as follows.

Firstly, the expression of matrices $G_l^{(\gamma)}$ is examined, by fairly assuming that at each stage:

- $E([s_k - \hat{s}_k][s_k^H - s_k^H]) \approx \rho(p_k) I_{N/2^{\gamma}}$

- $E(s_k \hat{s}_{k'}) \approx 0$ for $k \neq k'$
- $E(\hat{s}_k b^H) \approx 0$

where E is the expectation operator and ρ is a function of p_k, the bit error probability for the k^{th} block after its last equalization, depending on the constellation used.

Under these assumptions, it is possible to calculate the expression of the MMSE equalization matrix used at each stage:

$$G_k^{(\gamma)} = \begin{bmatrix} 0_{\frac{N}{2^\gamma} \times \frac{(k-1)N}{2^\gamma}} & I_{\frac{N}{2^\gamma}} & 0_{\frac{N}{2^\gamma} \times \frac{(2^\gamma - k)N}{2^\gamma}} \end{bmatrix} M^H \left(M D_k^{(\gamma)} M + \sigma^2 I_N \right)^{-1}$$

where $D_k^{(\gamma)}$ is the following block-diagonal matrix :

$$D_k^{(\gamma)} = diag\left(\rho(p_1) I_{\frac{N}{2^\gamma}} \quad \cdots \quad \rho(p_{k-1}) I_{\frac{N}{2\gamma}} \quad I_{\frac{N}{2^\gamma}} \quad \rho(p_{k+1}) I_{\frac{N}{2^\gamma}} \quad \cdots \quad \rho(p_{2^\gamma}) I_{\frac{N}{2^\gamma}} \right)$$

Simulations show that the terms $\rho(p_k)$ do not play an important role in overall performance, and thus can be neglected (replaced by 0), which greatly simplifies the calculus of the matrix products. It can be shown that in this case, when defining the $\frac{N}{2^\gamma} \times \frac{N}{2^\gamma}$ diagonal matrix:

$$\Delta_\gamma = diag\left\{ \frac{1}{\sigma^2 + \frac{1}{2^\gamma} \sum_{k=0}^{2^\gamma - 1} \left| h_{l+kN/2^\gamma} \right|^2} \right\}_{l=1}^{N/2^\gamma}$$

the following general result is obtained:

$$\begin{bmatrix} G_1^{(\gamma)} \\ \dots \\ G_{2^\gamma}^{(\gamma)} \end{bmatrix} = W \begin{bmatrix} \Delta_\gamma & 0 & 0 \\ 0 & \dots & 0 \\ 0 & 0 & \Delta_\gamma \end{bmatrix} H^*$$

Thus the product by $G_i^{(\gamma)}$ reduces simply to the product of y_i by a diagonal matrix depending on the channel coefficients (computed and stored once only) followed by the products of a subset of a Walsh-Hadamard matrix of size $\frac{N}{2^\gamma} \times N$. Therefore, the procedure detailed in the two previous equations results in a simple low arithmetical complexity way for performing the various MMSE equalizations steps. Instead of the expected heavy arithmetic complexity order of N^3 required by the $G_i^{(\gamma)}$ product, a much simpler complexity of order $2^\gamma N \log_2\left(\frac{N}{2^\gamma}\right)$ at each stage results.

The complexity of the multi-resolution decoding algorithm described above can be estimated as follows. The arithmetical simplifications due to the Walsh-Hadamard structure lead to quite a low complexity. At each stage γ of the algorithm, the complexity $C(\gamma,N)$ of one iteration (i.e., 2^γ calculus of y, of $G_i^{(\gamma)}y$ and decisions) can be overestimated:

$$C(\gamma,N) \approx N\left(2^\gamma\left(2\log_2\left(\frac{N}{2^\gamma}\right)+6\right)+2+\frac{3}{2^\gamma}\right)\times AddR + N\left(3\times 2^\gamma + 4 + \frac{5}{2^\gamma}\right)\times MulR + N \times Decision$$

where *AddR* is the complexity of an addition of two real values (assumed equal to that of a subtraction), *MulR* is the complexity of a multiplication, and *Decision*

is the complexity of a hard decision on a complex value (the choice of a symbol).

There follows an illustration of the performance improvement provided by the above-described multi-resolution decoding algorithm in the context of a 5.2GHz, 20MHz bandwidth, 64 carrier with 800ns guard time HIPERLAN/2 OFDM system using a QPSK constellation. Simulations were run using 2 channel profiles: (i) a perfect time interleaved BRAN 'E' channel model, and (ii) pure independent Rayleigh fadings in the frequency domain. The results in terms of bit error rate (BER) for uncoded scenarios as a function of the ratio $\frac{E_b}{N_0}$ (energy per bit / noise energy) are provided FIG. 4 and FIG. 5.

A clear improvement can be observed by applying the new decoding strategy compared with OFDM and MMSE SOFDM systems using a Walsh-Hadamard spreading sequence: for a target BER of 10^{-4}, more than 3dB is gained compared to MMSE SOFDM applying one or two iterations at the three first stages.

This means that for the same fixed BER and a given C/I (carrier to interference), 16QAM SOFDM with multiresolution decoding would have the same performance of a QPSK SOFDM MMSE transmission scheme while providing an enhancement of 4 times in bit rate. Such a significant improvement illustrates how improved decoding schemes for existing systems can

translate directly in greater system capacity under a given QoS constraint.

It will be understood that the multi-resolution decoding algorithm for OFDM-CDMA, spread OFDM single user systems described above provides the following advantages:
The following technical merits of the new multi-resolution decoding algorithm can be highlighted:

- Low arithmetical complexity compared to existing SIC BLAST techniques with same or better performance.
- Flexibility and scalability of the method (it is possible to adjust the number of iterations to be performed based on a performance/complexity tradeoff).
- Can be combined into all OFDM standards as a proprietary transmission mode (since it can be viewed as a simple extension of current OFDM systems).
- Yields a significant PER performance enhancement compared to classical OFDM and minimum mean square error (MMSE) SOFDM receivers (e.g., 3dB).